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Transistor Wien Bridge Oscillators for 20c/s to 300kc/s

J. H. Giddy
Wembley Laboratories, Associated Semiconductor Manufacturers Ltd.

A practical Wien bridge oscillator is described. The stabilised output is about 1·0V, and total distortion at 1·0kc/s is < 0·5%. Alternative component values are given for operation at fourteen different frequencies in the range 20c/s to 300kc/s. The measured frequencies are within ±3% of the theoretical values.

INTRODUCTION

The Wien bridge circuit (Fig. 1) has become very popular for use in R-C oscillators, because, in comparison with other configurations, the attenuation of the network at its zero-phase frequency is low (= 3 for a 'symmetrical' network; that is, one in which $R_1 = R_2$, and $C_1 = C_2$), and fewer components are used in the frequency-setting part of the circuit. The low attenuation allows a larger amount of negative feedback to be used, which gives a linear amplifier characteristic and, hence, low distortion in the output waveform.

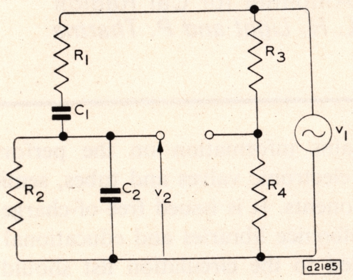


Fig. 1—Wien bridge

A disadvantage is that a two-stage amplifier must be used, since the network gives zero phase shift at its characteristic frequency; and a condition for oscillation is that the total loop phase shift must be a multiple of 360°.

This article discusses the problems that arise in designing a satisfactory circuit for a transistor Wien bridge oscillator. These problems include the effect on the operating frequency of a non-ideal amplifier, and the stabilisation of the amplitude of oscillation. A circuit is given which fulfils practical requirements. The frequency of operation can be selected in the range 20c/s to 300kc/s. The output voltage is stabilised at about 1·0V r.m.s.

THE WIEN BRIDGE

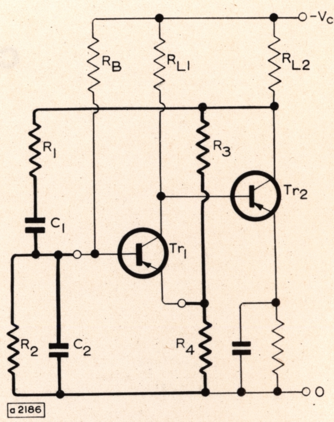


Fig. 2—Basic Wien bridge oscillator

The circuit of Fig. 1 may be used directly with a voltage amplifier as shown in Fig. 2. This circuit may also be considered as an amplifier whose gain is controlled by negative feedback via R_3 and R_4, and with positive feedback applied through a frequency-sensitive network—$R_1C_1R_2C_2$.

It is advantageous in some circumstances, as shown later, to consider also the dual network (Fig. 4).

It can be shown that the voltage transfer function, for the conventional network (Fig. 3), at the characteristic (zero phase shift) frequency, is:

$$\frac{v_1}{v_2} = 1 + \frac{R_1}{R_2} + \frac{C_1}{C_2} \qquad \ldots(1)$$

and the characteristic frequency, ω_0, is given by:

$$\omega_0 = \frac{1}{\sqrt{(R_1R_2C_1C_2)}}. \qquad \ldots(2)$$

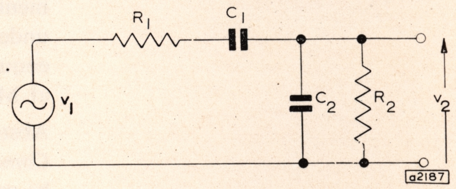


Fig. 3—Wien voltage network

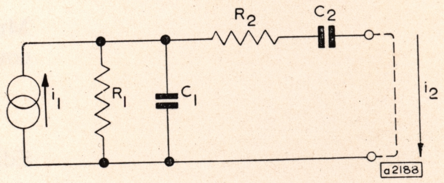


Fig. 4—Wien current network

It then follows, from the theory of dual networks, that the current transfer function for the dual network (Fig. 4) is identical with Eq(1). That is:

$$\frac{i_1}{i_2} = 1 + \frac{R_1}{R_2} + \frac{C_1}{C_2}$$

and

$$\omega_0 = \frac{1}{\sqrt{(R_1R_2C_1C_2)}}.$$

It may be shown that, for the symmetrical network ($R_1 = R_2$; $C_1 = C_2$), the rate of change of phase shift with frequency, at the characteristic frequency, is given by:

$$\left.\frac{\partial\phi}{\partial\omega}\right|_{\omega\,=\,\omega_0} = \frac{-2}{3\omega_0}. \qquad \ldots(3)$$

Oscillation will occur at a frequency where the overall loop phase shift (amplifier plus Wien network) is a multiple of 360°; so that if the amplifier has a non-zero phase shift, oscillation will occur at a frequency different from ω_0. That is, if there is a leading phase shift in the amplifier, the frequency will be higher than ω_0, and if a lagging phase shift exists, the frequency will be lower than ω_0.

It is also evident that the amplifier input and output impedances will affect the frequency of oscillation. In the conventional voltage system, the input resistance of the amplifier is in parallel with R_2 and the output resistance is in series with R_1. The oscillation frequency is thus given by:

$$\omega'_0 = \frac{1}{\sqrt{\left[C_1C_2(R_1+r_{out})\dfrac{R_2r_{in}}{R_2+r_{in}}\right]}} \qquad \ldots(4)$$

and for the dual system network of Fig. 4 and current amplifier:

$$\omega'_0 = \frac{1}{\sqrt{\left[C_1C_2\left(\dfrac{R_1r_{out}}{R_1+r_{out}}\right)(R_2+r_{in})\right]}}. \qquad \ldots(5)$$

(When a symmetrical network is used, Eqs(4) and (5) are identical.)

In practice, circuits are designed so that r_{in} and r_{out} have negligible effect on the frequency of operation.

AMPLIFIER DESIGN

For the reason outlined above, the amplifier must have very low phase shift over the required frequency range. Hence a directly-coupled two-stage amplifier is particularly suitable, since this eliminates one low-frequency time-constant, and reduces the number of large-value electrolytic capacitors required.

The amplifier suggested in Fig. 2 is not always satisfactory, because the biasing arrangement for Tr_1 will modify the frequency of oscillation, since R_B is effectively in parallel with R_2. It is desirable to connect the bias supply to a point of low impedance, so that the effective resistance of the bias supply may be low enough for good thermal stability of the d.c. conditions without seriously modifying the signal impedance at that point. For this reason the current form of the network with a directly-coupled current amplifier is preferable. A circuit of this type is given in Fig. 5.*

The bias for the first stage is derived from a bypassed resistor in the emitter circuit of Tr_2 through the 10kΩ resistor. This provides d.c. negative feedback which enhances the stability of the d.c. conditions and allows operation at ambient temperatures from 0° to 55°C with no significant variation of operating points.

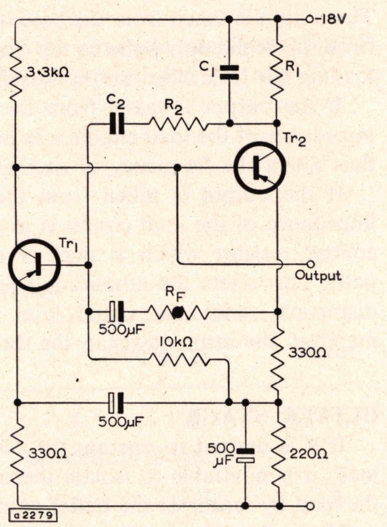


Fig. 5—Practical Wien bridge oscillator

The Wien network is $R_1C_1R_2C_2$, and R_1 becomes the load resistor for Tr_2.

The output resistance of the amplifier will be slightly larger than $1/h_{oe}$, and the input resistance will be approximately $R_F/(1+A_v)$, where A_v is the voltage gain, without feedback, from the base of Tr_1 to the emitter of Tr_2. In this particular case, $R_F \simeq 700\Omega$, and the input resistance is less than 3Ω.

It is necessary that R_1 (= R_2) is chosen so that it is 'large' compared with 3Ω but 'small' compared with $1/h_{oe}$ (which is of the order of 20 to 30kΩ).

The bias levels have been chosen such that if R_1 is greater than 1kΩ, then Tr_2 may be bottomed by the expected amplitude of oscillation. However, R_1 may be made smaller than 1kΩ with no difficulty.

AMPLITUDE STABILISATION

The amplitude of oscillation is made self-regulating by using a thermistor (S.T.C. type R53) as R_F. This is the simplest method and has only one disadvantage: it is sensitive to ambient temperature. The resistance of the thermistor depends on its temperature, which is a function of both the power generated in it by the signal and the power dissipated to the atmosphere. If the ambient temperature increases, less power is needed to keep the thermistor at a certain temperature, hence the signal level is reduced. Conversely, if the ambient temperature drops, more power is required to maintain the thermistor temperature constant, and the signal level rises.

This is not a great disadvantage in most applications, since the ambient temperature changes slowly. The thermistor is chosen to give approximately 1·0V output at the base of Tr_2 at an ambient temperature of 20°C.

---

*Cherry, E. M. 'An Engineering Approach to the Design of Transistor Feedback Amplifiers'. **The Radio and Electronic Engineer (Journ. Brit. I.R.E.)** Vol. 25, No. 2, February 1963, pp. 127 to 144.

The output is taken from the base of Tr_2 (collector of Tr_1) since this is the only point in the circuit at which moderate loading has little effect on the waveform of the signal.

If the output is taken from the collector of Tr_2, the impedance of the load circuit is in parallel with R_1, and it thus alters the frequency of oscillation.

If the output is taken from the emitter of Tr_2, the impedance of the load circuit is in parallel with the 330Ω emitter resistor which is part of the negative feedback path. This alters the amplitude of oscillation and causes distortion, since the signal level tends to increase to maintain the same power in the thermistor.

BUFFER STAGE

If it is desired to operate the oscillator into a varying load, it is advisable to isolate the oscillation circuit from the load by means of the buffer amplifier shown in Fig. 6.

Heavy loading on the output of the buffer amplifier will only reduce the amplitude of the signal. It will not alter the frequency or cause distortion. By using a potentiometer as the load of the buffer amplifier, a convenient amplitude control may be provided. The maximum output power of this amplifier is 0·5mW into a load of 1·0kΩ. If more power is required, the output may be fed into a conventional power amplifier. The maximum output voltage is about 1·0V r.m.s.

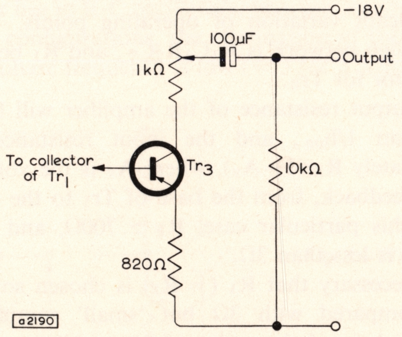


Fig. 6—Buffer amplifier

Suitable transistors for Tr_3 are the OC44, OC170 and GET889.

CHOICE OF TRANSISTORS

The limit to the maximum frequency which may be obtained with the circuit of Fig. 5 is set by the high-frequency phase shift in the amplifier. This depends primarily on the cut-off frequency of the transistors Tr_1 and Tr_2.

With audio transistors ($f_1 \simeq 1{\cdot}0$Mc/s) an amplifier phase shift of about 5° occurs at 30kc/s. R.F. transistors ($f_1 \simeq 15$Mc/s) produce the same phase shift at about 1·0Mc/s. So for an oscillator to operate at calculable frequencies up to a few hundred kc/s, it is necessary to use r.f. transistors.

The voltage, current, and power requirements for the transistors are quite modest. The approximate values for the circuit of Fig. 5 are given in Table 1.

TABLE 1

Transistor voltage, current, and power requirements

	V_{ce} (V)	I_c (mA)	P (mW)
Tr_1	4	4	16
Tr_2	9	8	72

It must be emphasised that the oscillator will oscillate at frequencies much higher than the limits given above for the respective types of transistor; but the frequency of operation depends more on the transistors themselves, and it becomes impossible to calculate the frequency accurately. Also, non-linear effects in the amplifier will become more noticeable as the loop gain of the negative feedback stabilised amplifier drops, thus increasing the waveform distortion.

Some suitable transistor types are suggested in Table 2.

TABLE 2

Recommended transistors

Maximum required oscillator frequency:	10kc/s	200kc/s
Tr_1	ACY22 OC70	OC44 OC170 GET889
Tr_2	ACY22	GET889

VALUES FOR WIEN NETWORK COMPONENTS

The circuit given in Fig. 5 has been designed on the basis that $R_1 = R_2$ and $C_1 = C_2$. Under these conditions the frequency is given by:

$$\omega_0 = \frac{1}{R_1C_1} = \frac{1}{R_2C_2}$$

or

$$f_0 = \frac{1}{2\pi R_1C_1} = \frac{1}{2\pi R_2C_2}.$$

The value of R_1 should be within the range 100Ω to 1·0kΩ to comply with the bias requirement mentioned earlier. If it is desired to make a tunable oscillator, then R_1 and R_2 must be varied together.

Representative component values, with theoretical and actual operating frequencies, are given in Table 3. The measurements were made with GET889 transistors and an 18V supply.

The calculated frequency is obtained from

$$f_c = \frac{1}{2\pi\sqrt{(R_1R_2C_1C_2)}}$$

using the measured values of R_1, R_2, C_1, and C_2. The accuracy of measurement of these values was 1%, thus giving an overall expected accuracy of ±2% for the calculated frequency.

It can be seen from the table that, except at the extreme ends of the frequency range, the overall accuracy is within ±3%.

PERFORMANCE

The performance of the practical oscillator (Fig. 5), with transistor type GET889 used for Tr_1 and Tr_2, is given in Table 4.

TABLE 3

Comparison of theoretical and actual frequencies of operation

R_1		**R_2**		**C_1**		**C_2**		**Frequency**		**Error**
nom. (kΩ)	meas. (kΩ)	nom. (kΩ)	meas. (kΩ)	nom. (μF)	meas. (μF)	nom. (μF)	meas. (μF)	f_c theoretical (kc/s)	f_m measured (kc/s)	$\frac{f_m - f_c}{f_c}$ (%)
1·0	0·960	1·0	0·961	8·0	8·35	8·0	8·22	0·020	0·0191	−4·4
1·0	0·960	1·0	0·961	1·0	0·92	1·0	0·96	0·176	0·171	−2·9
1·0	0·960	1·0	0·961	0·1	0·097	0·1	0·099	1·69	1·655	−2·37
1·0	0·960	1·0	0·961	0·03	0·0325	0·03	0·0305	5·27	5·19	−1·71
1·0	0·960	1·0	0·961	0·02	0·0184	0·02	0·0218	8·26	8·20	−0·73
1·0	0·960	1·0	0·961	0·01	0·0091	0·01	0·0088	18·5	18·39	−0·86
1·0	0·960	1·0	0·961	0·005	0·0058	0·005	0·0050	30·2	31·05	+2·84
0·10	0·100	0·10	0·1005	8·0	8·35	8·0	8·22	191·8	196·1	+2·24
0·10	0·100	0·10	0·1005	1·0	0·92	1·0	0·96	1·68	1·73	+2·98
0·10	0·100	0·10	0·1005	0·1	0·097	0·1	0·099	16·2	16·6	+2·47
0·10	0·100	0·10	0·1005	0·03	0·0325	0·03	0·0305	50·5	51·9	+2·77
0·10	0·100	0·10	0·1005	0·02	0·0184	0·02	0·0218	79·4	80·8	+1·64
0·10	0·100	0·10	0·1005	0·01	0·0091	0·01	0·0088	177·6	177·8	+0·11
0·10	0·100	0·10	0·1005	0·005	0·0058	0·005	0·0050	294·5	295·8	+0·44

TABLE 4

Performance

Supply voltage (V)	**Ambient temperature (°C)**	**Output (V)**	**D_{tot} at 1·0kc/s (%)**
18	25	0·92	< 0·5
18	35	0·86	—
18	55	0·73	—
15	25	—	< 1·0

Frequency-selective Amplifier for use with ORP13 Photoconductive Cell

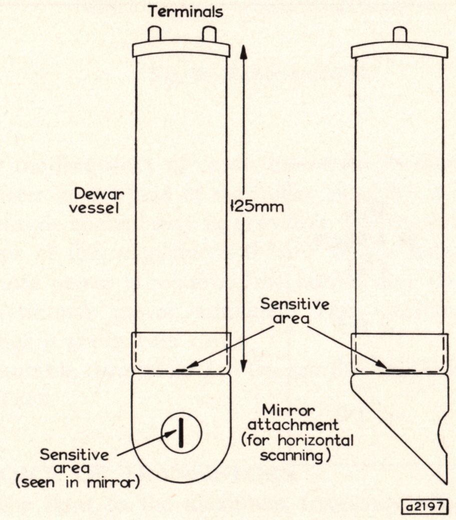
J. F. Miles
Mullard Semiconductor Measurement and Application Laboratory

A stable, linear, low-noise, five-stage transistorised a.c. amplifier, with associated frequency-selective network, demodulator, and reference circuit, is described. With 800c/s chopped radiation input to an ORP13 photoconductive cell, it provides a d.c. output indication on a 0–100μA meter. Radiation levels of the order of 10^{-11}W can be detected.

INTRODUCTION

A circuit for measuring the output from an ORP13 photoconductive cell is described. A narrow bandwidth system is employed, to minimise the noise contribution of both the cell and the amplifier. It is shown that radiation levels of the order of 10^{-11}W may be detected.

The Mullard ORP13 is an indium antimonide photoconductive cell, intended for the detection of chopped radiation in the range from visible to 5·4μm. The cell envelope is constructed in the form of a dewar vessel, so that the cell may be cooled to 77°K with liquid nitrogen. This increases the sensitivity and the signal-to-noise ratio. The generated noise of the cell is less than seven times the Johnson noise at 25°C. The amplifier is therefore required to have high gain with generated noise of the same order as Johnson noise.


Terminals
Dewar vessel
125mm
Sensitive area
Mirror attachment (for horizontal scanning)
Sensitive area (seen in mirror)
a2197

Mullard ORP13

If the radiation incident on the cell is chopped, the signal from the cell approximates to a square wave, the fundamental component of which may be taken to represent the signal. The noise may therefore be reduced by reducing the bandwidth of the amplifier.

This article describes a circuit working on the 'homodyne' system (Ref. 1), suggested by Avery and Bowes. The signal from the cell is applied to an amplifier with frequency-selective feedback, limiting the bandwidth to 50c/s centred on 800c/s. The output is applied to a phase-sensitive rectifier using, as reference, a signal derived from the same source as the radiation measured, and therefore locked to it in frequency and phase.

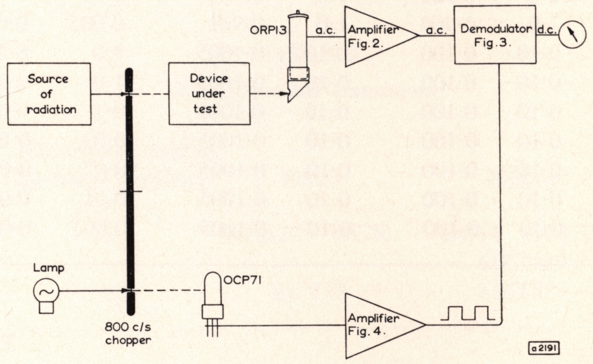


Fig. 1—Homodyne circuit

The output from the phase-sensitive rectifier is a d.c. current proportional to the signal, together with noise of frequencies up to 25c/s. The noise may now be reduced to any desired extent by smoothing. A smoothing time-constant of 0·3s gives an effective bandwidth, referred to the input, of 1c/s, centred on 800c/s.

A schematic diagram of the homodyne circuit (Ref. 2) is shown in Fig. 1.

A.C. AMPLIFIER

The a.c. amplifier (Fig. 2) consists of the ORP13 and its bias circuit, a low-noise input stage (Tr_1), and four further stages (Tr_2, Tr_3, Tr_4, and Tr_5).

First Stage

As the change in conductance of a photoconductive cell is proportional to the incident radiation, the normal

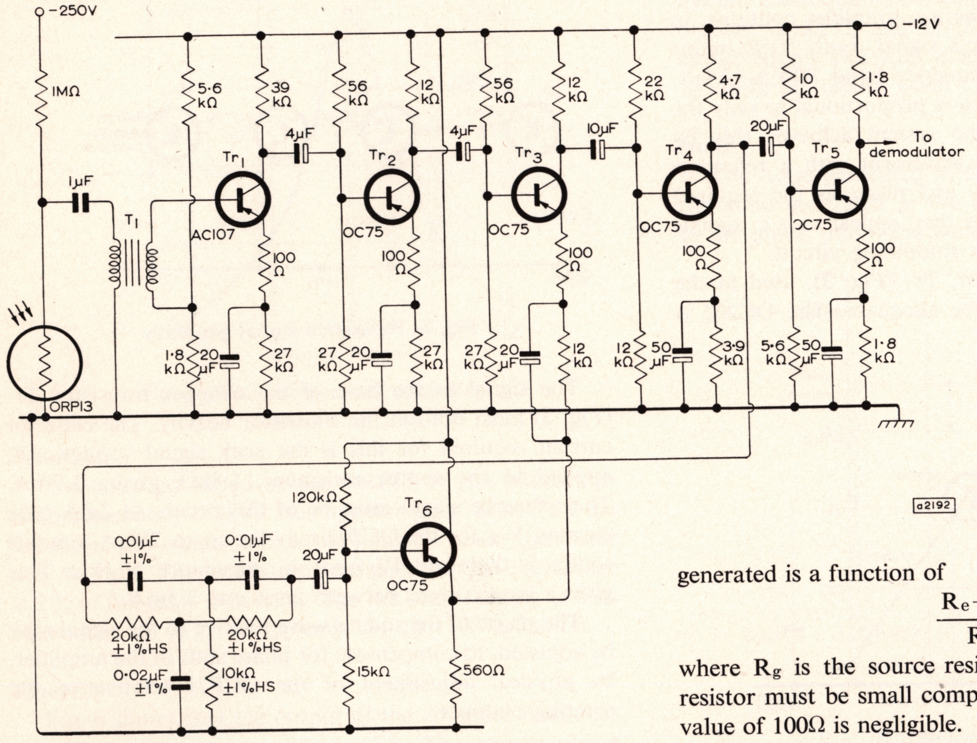


Fig. 2—
Photocell input stage, a.c. amplifier, and frequency-selective network

method of measurement is to pass a d.c. current through the cell from a high-impedance source and to measure the voltage developed. Since the radiation is chopped, the voltage developed is an 800c/s square wave.

As the maximum resistance of the cell is 60kΩ, and the optimum d.c. bias current is 250μA, the cell must be fed from a source of impedance much greater than 60kΩ—for example, 250V in series with 1MΩ, as shown in Fig. 2.

The first stage of the amplifier, Tr_1, must be designed for minimum generated noise. The most suitable transistor is therefore the AC107, which must be run at the minimum possible collector current. Account must be taken of the leakage current (which may be limited to 10μA at 40°C by a stabilising emitter resistor), the noise current (of the order of 0·1μA), and the signal current (of the same order). The collector current is therefore set at 100μA, giving a large safety factor.

At this current the source resistance for minimum noise factor is 2·5kΩ (Ref. 3); and, as the input impedance is of this order, approximate matching is obtained. The impedance-changing transformer T_1 must have a primary inductance greater than 35H at 800c/s. The Fortiphone P2329 is used, with a turns ratio of 3·3:1, matching the typical ORP13 resistance of 27kΩ into 2·5kΩ.

As the signal is fed in series with the bias, a low value of bias chain resistance can be used, and the factor of stability

$$\frac{R_b+R_e}{R_b+R_e(1+\beta)} \text{ reduces to } \frac{1\cdot 1}{1+\beta}$$

with a decoupled emitter resistor of 27kΩ. This reduces the effective I_{CEO} to nearly I_{CBO}, the figure assumed above. A feedback emitter resistor is added to stabilise the a.c. gain. Since, according to Coffey (Ref. 4) the noise generated is a function of

$$\frac{R_e+R_g}{R_g}$$

where R_g is the source resistance, this feedback emitter resistor must be small compared with 2·5kΩ. The chosen value of 100Ω is negligible.

Second and Subsequent Stages

The noise of the second stage, Tr_2, has also to be kept low. An OC75 is used, and the collector current is limited to the minimum. As I_{CBO} at 40°C is 50μA, and a factor of stability of 2/β is obtainable, I_{CEO} does not exceed 100μA. The signal and noise currents are small compared with the leakage, and a standing current of 150μA is therefore adequate in the collector. Feedback is again applied with an emitter resistor.

The other three stages (Tr_3, Tr_4, and Tr_5) are designed from similar considerations.

FREQUENCY-SELECTIVE NETWORK

Frequency-selective feedback is applied from the fourth stage, Tr_4, to the third stage, Tr_3, these stages being selected because the signal level is sufficiently high to swamp any noise generated in the network. The fifth stage, Tr_5, acts as a buffer stage between the network and the load.

A twin-T filter is used in the frequency-selective network, as it gives good stability and a sharp frequency response. A characteristic impedance of 7kΩ, the mean of the source and load impedances, is chosen, and an emitter follower, Tr_6, is used to terminate the network.

It has been found necessary to tune the filter accurately, as the large phase shifts caused by off-tuning make the circuit unstable. The components should be ±1% high stability, and the final tuning may be effected by padding the parallel components, 0·02μF and 10kΩ (Ref. 5).

As the filter is fed in parallel with the following stage of the amplifier, the input impedance of this stage must be high. The feedback emitter resistor is chosen to make the input impedance of the same order as that of the filter.

DEMODULATOR

The output signal from the a.c. amplifier contains, in addition to the desired 800c/s signal, both 50c/s mains pickup and noise in the frequency band 800c/s±25c/s. A d.c. signal is required which is proportional to only the 800c/s signal. The homodyne system achieves this by demodulation, chopping the waveform with a reference signal of the same frequency and phase as the required signal. The other components then appear as a.c. signals and may be filtered out in a smoothing circuit.

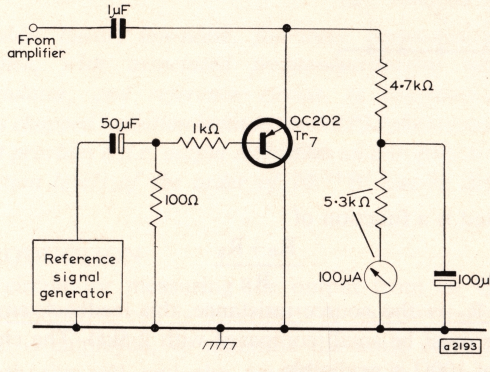


Fig. 3—Demodulator and meter circuit

A single chopper transistor, Tr_7 (Fig. 3), used in the inverted mode, is found to be adequate. The OC202 is suitable. In the 'off' state the voltage across the chopper transistor is

$$V_S . \frac{r_o}{R_S + r_o} + I_o R_S ,$$

where V_s is the signal, R_s the source resistance, r_o the 'off' impedance of the transistor, and I_o the offset current.

As R_s is low (1·8kΩ), I_oR_s is small; and, as r_o is of the order of 1MΩ,

$$V_S . \frac{r_o}{R_S + r_o} \simeq V_S$$

to within 1%.

In the 'on' state the voltage across the chopper transistor is

$$V_p + \frac{V_S r_c}{r_c + R_S} ,$$

where V_p is the offset voltage of the transistor and r_c is the 'on' resistance of the transistor. With the OC202 in inverted mode, r_c is approximately 20Ω, and V_p is less than 1mV. Thus

$$\frac{r_c}{r_c + R_S} = \frac{20}{20+1800} = \frac{1}{90} .$$

The on-to-off ratio is therefore 90, provided that V_s is large compared with 1mV. As the full output is 2V r.m.s., this condition is achieved.

DEMODULATOR REFERENCE CIRCUIT

The reference signal to switch the chopper transistor must be derived through the same rotating chopper as the signal, and this may conveniently be achieved by the use of an OCP71, as in Fig. 1. The circuit used depends on the intensity of light available, but will be of the type shown in Fig. 4.

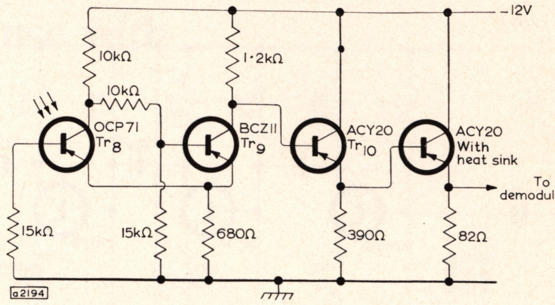


Fig. 4—Reference signal generator

The signal at the base of the chopper transistor Tr_7 (Fig. 3) must bottom the transistor heavily. The collector current required for this is the peak signal voltage, 3V, applied to the source resistance, 1·8kΩ, giving 1·7mA. The reference signal must be of this order. At 2mA base current, V_{be} for the OC202 may be up to 2V. A suitable source is therefore 4V from an impedance of 1kΩ. This gives a base current between 2mA and 3·5mA.

The phase of the square wave relative to the signal may be adjusted, to compensate for phase shift in the amplifier, by physical adjustment of the OCP71 relative to the rotating chopper.

METER CIRCUIT

The output from the demodulator, if the phase of the reference signal is adjusted to be in phase with the signal, is a half-wave rectified 800c/s sinewave, together with pickup and noise, chopped at 800c/s. The source impedance, while the signal is present, is 1·8kΩ. The readout circuit must therefore have an impedance large compared with 1·8kΩ (for example, 10kΩ) and a time-constant sufficiently long to smooth the low-frequency noise mentioned in the introduction. A simple microammeter, 0 to 100μA, with sufficient series resistance to make its total resistance 10kΩ, may be adequate. If the time-constant of the meter alone is not sufficiently long, smoothing may be added, as shown in Fig. 3.

PROTOTYPE AMPLIFIER

An amplifier has been built, using the circuits of Figs. 2, 3, and 4. It has been constructed on Veroboard. Normal precautions have been taken in the layout to avoid pickup, and the transformer is mounted inside an earthed box, all leads to it being screened. Special care in earthing is required to avoid spurious input signals being picked up from the reference signal.

Frequency Response

As the bandwidth of the amplifier is controlled by the frequency-selective feedback, it may be varied by changing the amount of feedback. It is found, however, that if additional amplification is inserted in the feedback loop, then the circuit becomes unstable, and complex phase correction becomes necessary.

With the amplifier shown in Fig. 2, the frequency response of Fig. 5 is obtained, with a bandwidth at the 3dB points of about 50c/s.

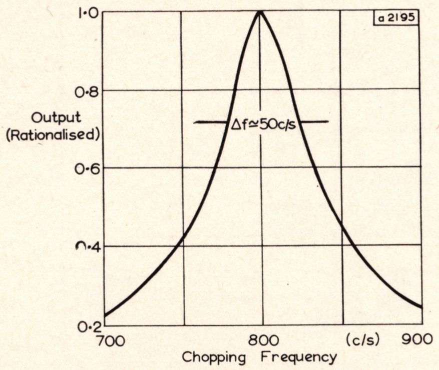


Fig. 5—Overall frequency response

Sensitivity

The noise generated by the ORP13 in 1c/s bandwidth is approximately 0·2μV. This appears as a 'dither' on the needle of the meter, and is the limit to which the meter may be read. The amplifier is designed so that this dither is 2% of full-scale deflection. The full-scale reading is then equivalent to 7μV, or typically 5×10^{-10}W radiation.

Linearity

Fig. 6 shows that the output of the amplifier is a linear function of the input to within ±1%. The zero is subject to an error due to interaction between the switching signal and the input. By careful earthing and screening this may be reduced to 2% of full scale.

Variation with Temperature

The first and second stages are designed to operate up to 40°C. Above this temperature, leakage current bottoms the transistors. The negative feedback holds the gain to within ±2% up to this temperature. As d.c. amplifiers are avoided, the zero level remains constant.

Variation with D.C. Supply Voltage

A change of 2% in d.c. supply voltage causes a change of 5% in the gain of the amplifier. A stabilised supply is therefore essential. This can be obtained with two zener diodes, OAZ212 and OAZ211, in series, shunting the circuit supply rail with a series resistor adjusted to give 10mA in the diodes when 15mA flows in the circuit.

Noise

The noise generated in the circuit is slightly less than the noise generated in the cell, which is 1·3μV, referred to the input, into 50c/s bandwidth.

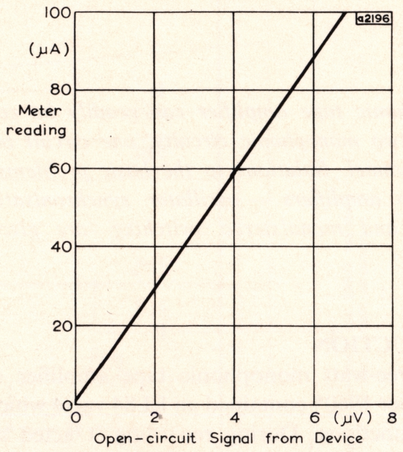


Fig. 6—Overall linearity

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Stereophonic Tape Recorders

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A stereophonic tape amplifier can readily be constructed based on two monophonic circuits, one circuit being used in each channel. Solutions to the basic problems of combining two amplifiers — oscillator synchronisation, level indication and mono/stereo switching—are given in this article.

INTRODUCTION

The three-watt monophonic tape amplifier described in September 1960* contained an EL84 valve which carried out two functions. During recording it acted as a bias and erase oscillator, and on playback the EL84 was used in the output stage to drive a loudspeaker.

Two complete power amplifiers are necessary for a stereophonic playback system. These can conveniently take the form of two of these three-watt tape amplifiers.

A common oscillator may be used for a stereophonic recording system, but there is a disadvantage in that complex switching of heads and dummy loads is introduced for changing from monophonic to stereophonic recording. This switching can also result in the heads being left in a magnetised state. To avoid this switching, two separate monophonic amplifiers can be used, each permanently connected to one set of heads.

There remain three design problems: synchronisation of the two oscillators, recording level indication and power supply arrangements. Each of these design problems is dealt with in detail. A balance control will also be required which may be readily fitted next to the existing volume control.

SYNCHRONISATION OF THE TWO OSCILLATORS

Both amplifiers are used simultaneously when stereophonic recording and playback are taking place. It is essential that the oscillators in the amplifiers are synchronised while recording, otherwise beating may occur between the two oscillators, causing unwanted signals to appear in the recording. The oscillators are synchronised by connecting a small capacitor (47pF) between the control grids of the two EL84 valves and by tuning each oscillator coil to approximately the same frequency, whereupon the oscillators lock. The tuning is carried out by connecting the output from each oscillator to the X and Y plates of an oscilloscope. The slug in each oscillator coil is adjusted and when the frequencies of the two oscillators are the same, an elliptical Lissajous figure appears. A valve voltmeter is used to ensure that the outputs of the oscillators are approximately equal. It is a simple operation to synchronise the oscillator frequencies by adjusting the slugs in the oscillator coils.

MODULATION LEVEL INDICATOR

There are three possible methods of indicating the modulation level while recording. The first method consists of using one indicator independently in each amplifier, but there are disadvantages in that it is inconvenient to view two indicators simultaneously. It is also more expensive to have two separate level indicators. A second method is to use one indicator which can be switched to the recording output of either amplifier. However, there is an obvious disadvantage when using this method in that one amplifier may be overloaded while the indicator is switched to the second amplifier, without the operator knowing that overloading is occurring. The third method of indication, which is adopted in the circuit of Fig. 1, is to make the indicator respond to whichever of the two signals is the greater at any one time. This is achieved by the use of an EB91 double diode connected as shown in the circuit.

The voltage-level indicator used is the EM87. An 82kΩ resistor may be used as the target load, but for accurate indication of the recording level a preset resistor is recommended. The preset resistor is adjusted so that the indicator closes when the input voltage to the amplifier produces 9V at 1kc/s at the second anode of the ECC83. The figure of 9V corresponds to a current of 80μA through the recording head.

POWER SUPPLY

When both amplifiers are used for stereophonic recording or playback, the h.t. current is 110mA at 300V and the l.t. current is 3·06A at 6·3V. One method of monophonic recording, when only one amplifier is used, is to switch off the h.t. to the amplifier which is connected to the lower track, since the accepted standard for monophonic recording is to use the upper track. (It is essential, therefore, that the h.t. connection to the EM87 indicator is made from the upper-track amplifier.) To avoid a large change of line voltage, this switching may be carried out in two ways. Firstly, it can be performed by switching a load resistor in place of the lower track amplifier (this would be a 6kΩ, 15W wirewound type). Secondly, it can

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*Dallosso, P. F. 'Three-Watt Tape Amplifier Circuit for Modern Tape Decks', **Mullard Technical Communications,** Vol. 5, No. 45, September 1960, pp. 191 to 196.

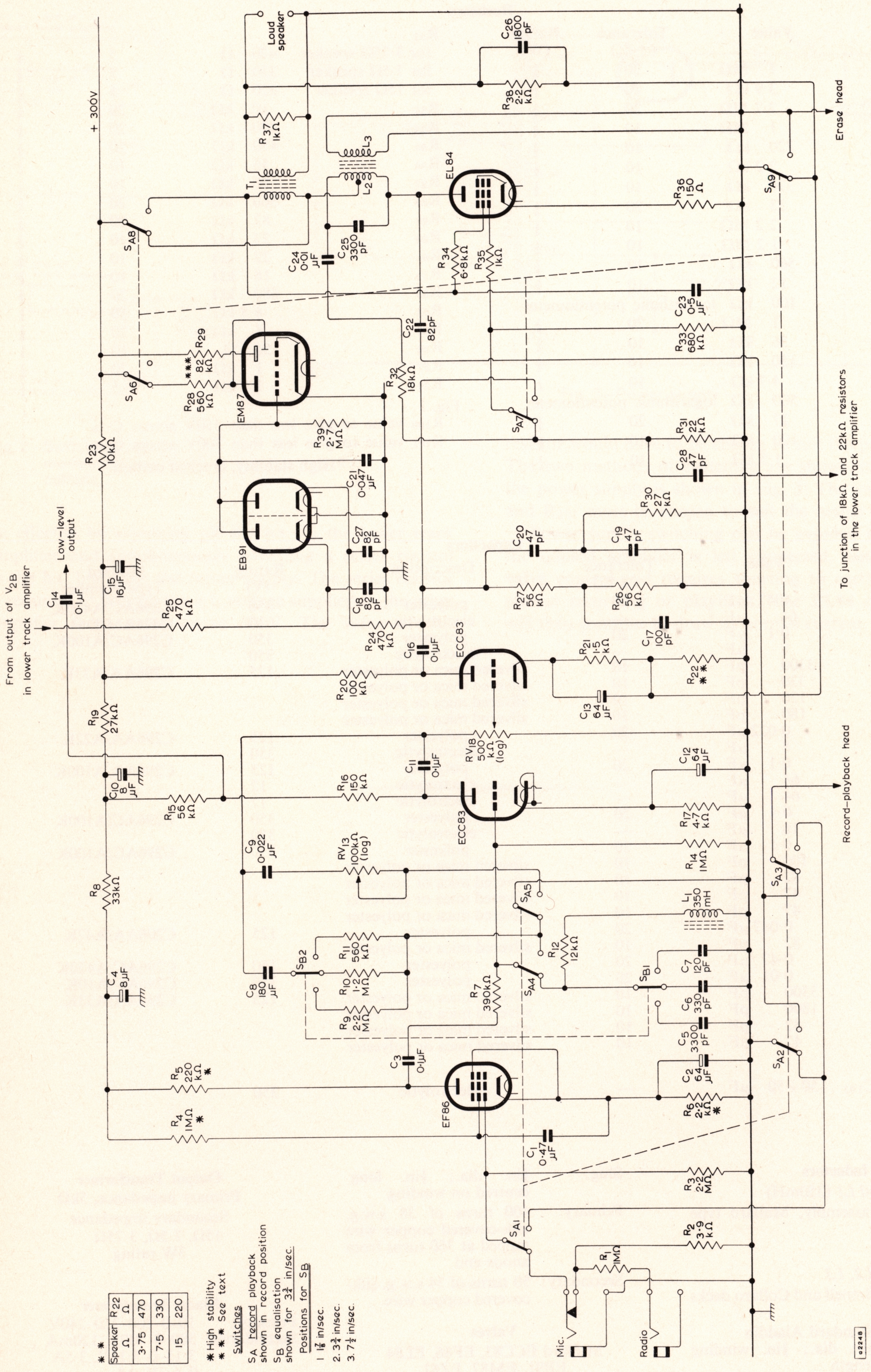
+300V
Loud speaker
Erase head
Record–playback head
From output of V2B in lower track amplifier
Low-level output
To junction of 18kΩ and 22kΩ resistors in the lower track amplifier
EL84
EM87
EB91
ECC83
EF86
Mic.
Radio
** Speaker Ω / R22 Ω: 3·75 470; 7·5 330; 15 220
* High stability
*** See text
Switches
SA record playback shown in record position
SB equalisation shown for 3¾ in/sec
Positions for SB
1. 1⅞ in/sec.
2. 3¾ in/sec.
3. 7½ in/sec.

Fig. 1—Three-watt tape amplifier
Full components list on page 112

Resistors

Circuit ref.	Value	Tolerance (±%)	Rating (W)
R_1	1 MΩ	20	¼
R_2	3·9 kΩ	20	¼
R_3	2·2 MΩ	20	¼
[1]R_4	1 MΩ	10	½
[1]R_5	220 kΩ	10	½
[1]R_6	2·2 kΩ	10	½
R_7	390 kΩ	10	¼
R_8	33 kΩ	20	¼
R_9	2·2 MΩ	10	¼
R_{10}	1·2 MΩ	10	¼
R_{11}	560 kΩ	10	¼
R_{12}	12 kΩ	10	¼
RV_{13}	100 kΩ	logarithmic potentiometer	
R_{14}	1 MΩ	20	¼
R_{15}	56 kΩ	10	¼
R_{16}	150 kΩ	10	¼
R_{17}	4·7 kΩ	10	¼
RV_{18}	500 kΩ	logarithmic potentiometer	
R_{19}	27 kΩ	20	¼
R_{20}	100 kΩ	10	¼
R_{21}	1·5 kΩ	10	¼
R_{22}			
for 3·75Ω speaker	470 Ω	5	¼
for 7·5Ω speaker	330 Ω	5	¼
for 15Ω speaker	220 Ω	5	¼
R_{23}	10 kΩ	20	¼
R_{24}	470 kΩ	20	¼
R_{25}	470 kΩ	20	¼
R_{26}	56 kΩ	10	¼
R_{27}	56 kΩ	10	¼
R_{28}	560 kΩ	20	¼
R_{29}	82 kΩ	20	¼
R_{30}	27 kΩ	10	¼
R_{31}	22 kΩ	10	¼
R_{32}	18 kΩ	10	¼
R_{33}	680 kΩ	20	¼
R_{34}	6·8 kΩ	10	1
R_{35}	1 kΩ	20	¼
R_{36}	150 Ω	10	1
R_{37}	1 kΩ	20	¼
R_{38}	2·2 kΩ	5	¼

Fig. 3:

R_{100} value to give less than 350V across C_{100}

R_{101} value to give less than 300V across C_{101}

[1]High stability, cracked carbon

Capacitors

Circuit ref.	Value	Tolerance (±%)	Description	Rating (V)	Mullard Type Number
C_1	0·47 μF	20	polyester	350	C296AC/A470K
C_2	64 μF	—	electrolytic	10	C426AM/D64
C_3	0·1 μF	20	polyester	350	C296AC/A100K
C_4	8 μF	—	electrolytic	350	
C_5	3300 pF	10	silvered mica or polyester	125	C296AA/A33K
C_6	330 pF	10	silvered mica or polyester		
C_7	120 pF	10	silvered mica or polyester		
C_8	180 pF	10	silvered mica or polyester		
C_9	0·022 μF	20	polyester	125	C296AA/A22K
C_{10}	8 μF	—	electrolytic	350	
C_{11}	0·1 μF	20	polyester	125	C296AA/A100K
C_{12}	64 μF	—	electrolytic	12	
C_{13}	64 μF	—	electrolytic	12	
C_{14}	0·1 μF	20	polyester	350	C296AC/A100K
C_{15}	16 μF	—	electrolytic	350	
C_{16}	0·1 μF	20	polyester	350	C296AC/A100K
C_{17}	100 pF	10	silvered mica or polyester		
C_{18}	82 pF	10	silvered mica or polyester		
C_{19}	47 pF	10	silvered mica or polyester		
C_{20}	47 pF	10	silvered mica or polyester		
C_{21}	0·047 μF	20	polyester	125	C296AA/A47K
C_{22}	82 pF	10	silvered mica or polyester		
C_{23}	0·47 μF	20	polyester	350	C296AC/A470K
C_{24}	0·01 μF	20	polyester	350	C296AC/A10K
C_{25}	3300 pF	10	silvered mica or polyester	125	C296AA/A33K
C_{26}	1800 pF	10	silvered mica or polyester		
C_{27}	82 pF	10	silvered mica or polyester		
C_{28}	47 pF	10	silvered mica or polyester		
Fig. 3					
C_{100}, C_{101}	50+50 μF	—	electrolytic	350	

Inductors

Equalisation Coil L1 (350mH)

Wound coil assembly, Mullard type WF1932

Oscillator Coil L2, L3

(suitable for Brenell and Collaro decks only)

Former:	Standard Aladdin ½in. dia.; ⅜in. winding length
Slug:	⅜in. dia.; ½in. long centred on winding
Primary:	400 turns of 38 s.w.g. silk-covered copper wire tapped at 360 turns from anode end
Secondary:	50 turns of 34 s.w.g. silk-covered copper wire

Valves

Mullard ECC83, EF86, EL84
EB91, EM87, EZ81

Output Transformer

Primary impedance, 5kΩ

Secondary impedance
15Ω, 7·5Ω, 3·75Ω
3W rating

Mains Transformer

Primary:	10–0–200–220–240V
Secondaries:	300–0–300V, 120mA 3·15–0–3·15V, 3·5A 0–6·3V, 1A

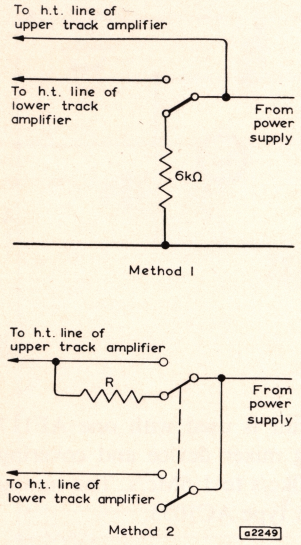
To h.t. line of upper track amplifier
To h.t. line of lower track amplifier
From power supply
6kΩ
Method 1
To h.t. line of upper track amplifier
R
From power supply
To h.t. line of lower track amplifier
Method 2
a2249

Fig. 2—Switching network for mono and stereo applications

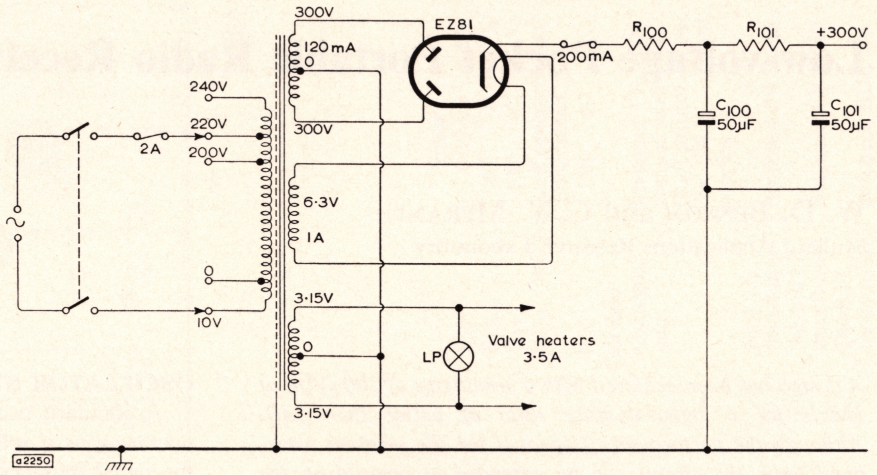

Fig. 3—Power supply

be effected by switching off the h.t. to the lower track amplifier and simultaneously switching a series resistor in the h.t. line from the power supply. The series resistor is required to prevent the h.t. voltage supplied to the other amplifier from rising above 300V. The value of this resistor would have to be chosen to suit the power supply. The two methods of switching are shown in Fig. 2 and the power supply is shown in Fig. 3. The values of R_{100} and R_{101} must be chosen to suit the mains transformer.

Power supply switching can be avoided if the h.t. of the unused amplifier is left connected, but the output valve control-grid requires shorting to earth for monophonic recording or playback. The EL84 output valve will then continue to draw its normal current.

Low-voltage Pocket Portable Radio Receiver

W. D. Benson and C. V. Merani
Mullard Applications Research Laboratory

A design has been achieved with a sensitivity of 100µV/m at 1Mc/s for a signal-to-noise ratio of better than 5dB. Although the performance is quoted for the medium waveband only, the circuit can be extended to operate on the long waveband as well.

With 3·0V operation the limitation of voltage swing reduced the gain of the r.f. and i.f. stages to 90dB as opposed to 100dB on the 4·5V design. Also there is insufficient power output from an a.f. amplifier working on a 3V supply.

INTRODUCTION

Preliminary calculations indicate that it is possible to obtain longer battery life in a portable receiver without significant loss of performance, if the circuit is designed to operate at as low a battery voltage as possible. If two or three 1·5V miniature cells are used in place of the more usual 9·0V battery, running costs per hour will be much reduced. Investigations were carried out to examine the working of AF117 transistors in the r.f. and i.f. stages of a small personal receiver operating from a 4·5V battery.

A satisfactory design for these stages has been achieved, which gives an overall gain of nearly 100dB when working from a supply voltage of 4·5V. When these stages form a complete receiver with a three-stage a.f. amplifier, the sensitivity of the receiver at 1Mc/s is 100µV/m for a signal-to-noise ratio of better than 5dB.

A 6dB-bandwidth of ±2·8kc/s and a ±9kc/s attenuation of better than 24dB are achieved. The receiver is capable of operating over an input-signal range of 63dB with automatic gain control over 40dB of the range. The maximum input signal which the receiver can handle is 200mV/m at 30% modulation. The r.f. stages operate satisfactorily even when the supply voltage falls to 3·0V, whilst the oscillator frequency stability during this voltage-change is 750c/s per V in the worst condition. The battery drain of the r.f. and i.f. stages is 4mA.

CIRCUIT DESIGN

The design methods adopted are mainly similar to those discussed in an earlier article*. Three AF117 transistors are used in the r.f. and i.f. sections. The a.f. stages consist of an AC127 and LFK3 audio package. (OC81D, AC127 and OC81 transistors). The complete circuit diagram is shown in Fig. 1.

OSCILLATOR STAGE

A standard oscillator circuit is used with one AF117 operating as a self-oscillating mixer device and covering the signal frequency band 550kc/s to 1·6Mc/s. The tuning gang is a Mullard component type AC1033.

I.F. STAGES

Two AF117 transistors are used in the i.f. stages together with three single-tuned i.f. transformers.

Stability

The stability condition for the i.f. stages is given by the expression

$$G_i G_o = S\frac{y_{fe}y_{re}}{B},$$

where G_i, G_o are the total input and output conductances of the base and collector circuits respectively. The factor B is a constant whose value depends upon the types of i.f. transformers used. For single-tuned stages, B has a value of 2. The terms y_{fe} and y_{re} are the forward and reverse transfer admittance of the transistor respectively and S is the stability factor. The reverse transfer admittance y_{re} is the susceptance of the maximum value of the feedback capacitor c_{re}. The feedback capacitance value is taken as 2·9pF at 470kc/s with a collector voltage V_c of 3V and an emitter current I_e of 1mA. The forward transfer admittance y_{fe} is 37mA/V. Therefore, the stability condition used is:

$$G_i G_o = \frac{S}{2}(37\times10^{-3})(2\pi\times470\times10^{3}\times2{\cdot}9\times10^{-12}),$$

that is,

$$G_i G_o = \frac{S}{6{\cdot}3\times10^{6}}. \qquad \ldots(1)$$

TABLE 1
Comparison of Gain and Gain Spreads for Different Stability Factors

Stability Factor S	Loss of gain (compared with that when S is 4) (dB)	Spreads due to G_{ie} alone (dB)
4	—	±2
6	1·8	±1·5
10	4·0	±1·0

A stability factor of 10 (see Table 1) is assumed for the second i.f. stage to reduce the value of G_i and thus minimise the spreads due to changes in g_{ie} of the transistor. The comparative loss of gain and advantage in spreads (cal-

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*Beckley, J. C. and Benson, W. B. 'Medium and Long-Wave Portable Receiver using Mullard AF117 Transistors'. **Mullard Technical Communications**, Vol. 6, No. 53, December 1961, pp. 114 to 121.

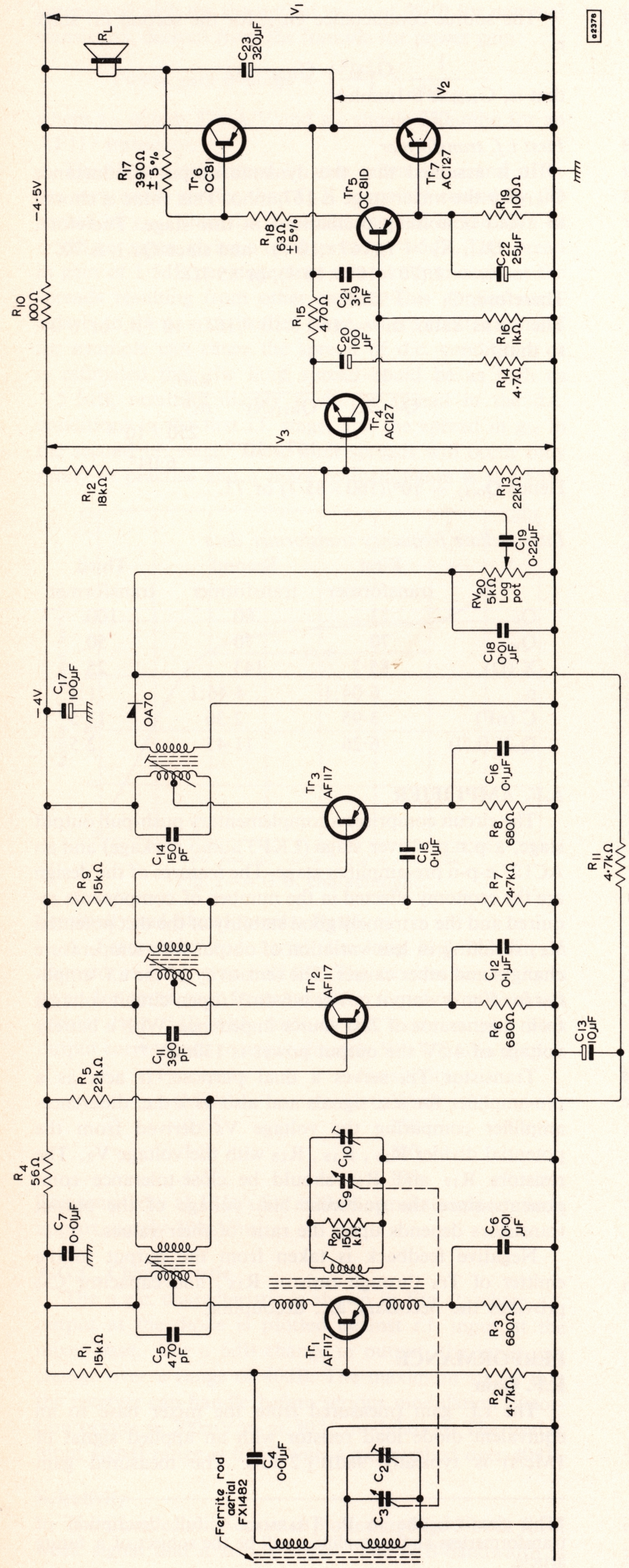

Fig. 1—Complete receiver

COMPONENTS

Capacitors

	Value	Suitable Mullard Type		Value	Suitable Mullard Type
C_2	*	—	C_{14}	150 pF	—
C_3	*	—	C_{15}	0·1 µF	C280AA/P100K
C_4	0·01 µF	C280AA/P10K	C_{16}	0·1 µF	C280AA/P100K
C_5	470 pF	—	C_{17}	100 µF	C426AM/B100
C_6	0·01 µF	C280AA/P10K	C_{18}	0·01 µF	C280AA/P10K
C_7	0·01 µF	C280AA/P10K	C_{19}	0·22 µF	C296AA/A220K
C_9	*	—	C_{20}	100 µF	C426AE/B100
C_{10}	*	—	C_{21}	3·9 nF	—
C_{11}	390 pF	—	C_{22}	25 µF	C426AE/B25
C_{12}	0·1 µF	C280AA/P100K	C_{23}	320 µF	C426AE/D320
C_{13}	10 µF	C426AM/E10			

*Part of Mullard AC1033 tuning gang

Resistors

All resistors are Mullard $\frac{1}{8}$W B8.305 series.

	Value		Value
R_1	15 kΩ	R_{17}	0·39 kΩ ±5%
R_2	4·7 kΩ	R_{18}	63 Ω ±5%
R_3	680 Ω	R_{19}	100 Ω
R_4	56 Ω	RV_{20}	5kΩ Mullard potentiometer E088 series.
R_5	22 kΩ	R_{21}	150 kΩ
R_6	680 Ω		
R_7	4·7 kΩ		
R_8	680 Ω		
R_9	15 kΩ		
R_{10}	100 Ω		
R_{11}	4·7 kΩ		
R_{12}	18 kΩ		
R_{13}	22 kΩ		
R_{14}	4·7 Ω		
R_{15}	0·47 kΩ		
R_{16}	1 kΩ		

Coil data

Oscillator coil:	Brayhead FP 50234 damped with 150kΩ (R_{21}) resistor across the M.W. tuned winding.
First i.f. transformer:	Brayhead FP 50245.
Second i.f. transformer:	Brayhead FP 50247.
Third i.f. transformer:	Brayhead BDL 546.

culated) are presented in the Table for stability factors of 4, 6 and 10.

For the first i.f. stage, S is taken as 4.

Bandwidth

The approximate performance desired was a 6dB bandwidth of ±3kc/s together with 23dB attenuation at 9kc/s. To achieve this, the working Q factor, Q_w, must be 70 for the first and second i.f. transformers and 50 for the third.

Transformer Data

Third i.f. transformer

The reactance X_3 of the tuning capacitor is given by:

$$X_3 = n_3^2 R_L \frac{Q_{o(3)} - Q_{w(3)}}{Q_{o(3)} \, . \, Q_{w(3)}} .$$

The undamped Q-factor, $Q_{o(3)}$, is assumed to be equal to 100. The equivalent diode damping resistance R_L is 2·5kΩ for the 5kΩ potentiometer used as the diode load. To obtain the requisite a.g.c. operation, the turns ratio n_3 for the third i.f. transformer is unity. Therefore,

$$X_3 = (1)^2 \times 2{\cdot}5 \times 10^3 \times \frac{50}{100 \times 50} = 25\Omega .$$

The tuning capacitance C_3 is given by:

$$C_3 = \frac{1}{2\pi \, f \, X_3} ,$$

so that, at a frequency of 470kc/s,

$$C_3 = \frac{1}{2\pi 470 \times 10^3 \times 25} \text{ F}, \; C_3 = 13{\cdot}5\text{nF}.$$

The output conductance $G_{o(3)}$, which is given by the reciprocal of Q_wX, is equal to 1/50×1/25, or 800μmho.

From Eq(1):

$$G_{i(3)} = \frac{S}{6{\cdot}3 \times 10^6 \times G_{o(3)}}$$
$$= \frac{10}{6{\cdot}3 \times 10^6 \times 0{\cdot}0008} .$$

Therefore, for a stability factor S of 10, $G_{i(3)}$ is 1/504mho or 1·98mmho.

Second i.f. transformer

It is assumed that the dynamic conductance $G_{o(2)}$ is 100μmho and $Q_{w(2)}$ is 70 for this stage. Therefore the reactance X_2 of the tuning capacitor is $1(70 \times 100 \times 10^{-6})$, or 143Ω, and the capacitance C_2 is 2·36nF.

The turns ratio n_2 is given by the equation

$$n_2 = \left\{ \frac{G_{i(3)}}{G_{o(2)}} \right\}^{\frac{1}{2}} ,$$

so that n_2 is 4·46:

The unloaded Q-factor $Q_{o(2)}$ is given by the expression

$$\frac{1}{Q_{o(2)}X_2} = \frac{1}{Q_{w(2)}X_2} - \frac{g_{ie}}{n_2^2} .$$

At 1mA, g_{ie} is 250μmho. Therefore,

$$\frac{1}{Q_{o(2)}X_2} = 10^{-4} - \left(\frac{250 \times 10^{-6}}{4{\cdot}46^2} \right)$$

so that $Q_{o(2)}X_2$ is 11·45kΩ and $Q_{o(2)}$ is approximately 80. From Eq(1), with a stability factor of 4,

$$G_{i(2)} = \frac{4}{6{\cdot}3 \times 10^6 \times 10^{-4}}$$

that is, $G_{i(2)}$ is 6·35mmho.

The source conductance $G_{s(2)}$ for the second i.f. stage is

$$G_{s(2)} = G_{i(2)} - g_{ie}$$

that is, $G_{s(2)}$ is 6·1mmho.

First i.f. transformer

It is assumed that the dynamic output conductance $G_{o(1)}$ for the mixer stage is 167μmho. This value is chosen to avoid bottoming oscillations in this stage. Therefore,

$Q_{w(1)}X_1 = 1/(167 \times 10^{-6})$ and since $Q_{w(1)}$ is 70,

$X_1 = 1/(70 \times 167 \times 10^{-6})$, or 85·7Ω.

Therefore, $C_1 = 3{\cdot}95$nF.

The turns ratio $n_1 = (6{\cdot}1 \times 10^{-3}/167 \times 10^{-6})^{\frac{1}{2}}$, or 6·04: so that,

$$\frac{1}{Q_{o(1)}X_1} = \frac{1}{Q_{w(1)}X_1} - \frac{g_{ie}}{n_1^2}$$
$$= 167 \times 10^{-6} - \frac{250 \times 10^{-6}}{6{\cdot}04^2} .$$

Hence $Q_{o(1)} = 10^6/(160 \times 85{\cdot}7)$ or 73.

Intermediate frequency transformer data

	First transformer	Second transformer	Third transformer
Q_o	73	80	100
Q_w	70	70	50
X (Ω)	85·7	143	25
n	6·04:1	4·46:1	1:1
C (nF)	3·95	2·36	13·5
Q_oX (kΩ)	6·26	11·45	2·5

A.F. AMPLIFIER

The circuit comprises a complementary push-pull output stage, a p.n.p. driver stage (LKF3 audio package) and an AC127 n-p-n pre-amplifier stage. The features of the design are the economy effected in the number of components required and the extremely good stability of the d.c. potential V_2, resulting in less variation of output with temperature changes and other causes. The sensitivity of the a.f. amplifier for 50mW output power is 6·4mV (open-circuit voltage) from a generator of 2kΩ source impedance. With a battery voltage of 4·5V the output power is 130mW.*

Transistor Tr_4 serves a dual purpose. It acts as a pre-amplifier for a.c. signals and also as a d.c. difference-amplifier comparing the voltage V_3 derived from the potential divider R_{10}, R_{12}, R_{13} with the voltage V_2. The resistors R_{17} and R_{18} should be close-tolerance components since the quiescent bias voltage of the output transistors depends upon the ratio of their values.

Negative feedback is taken from the output to the emitter of Tr_4 through resistor R_{15}, and capacitor C_{20} provides the necessary a.c. decoupling.

PERFORMANCE

R.F. Gain

The r.f. gain (measured from the mixer base to an equivalent diode load resistor with an applied signal of 1Mc/s) is typically 98dB±2·5dB. The measured gain agrees well with the theoretical value of 99dB for nominal transistors, derived from the formula for power gain:

$$\text{Power gain of stage} = \left(\frac{y_{fe}}{G_o} \right)^2 \times \frac{1}{n^2 R_L g_{ie}} ,$$

where y_{fe} equals 37mA/V and g_{ie} equals 250μmho for an AF117 transistor.

The theoretical spread in gain when $1/g_{oe}$ and n^2R_L are very much greater than Q_oX is mainly caused by the spread of y_{fe}. Thus the theoretical spread is $(38/32)^2$ or 1·5dB per stage, giving a total spread of 4·5dB. The spread in gain of 1·5dB per stage does not take into account the spreads resulting from other causes. This applies particularly to the gain controlled stage, where resistor spreads for example can cause the spread of d.c. conditions. It is estimated that the limit spread could be as high as ±5·5dB, resulting mainly from the spread in the d.c. conditions in the first i.f. stage, and the spread in g_{ie} in the second i.f. stage. These limit spreads will occur only rarely in practice and a more realistic figure is ±3·5dB.

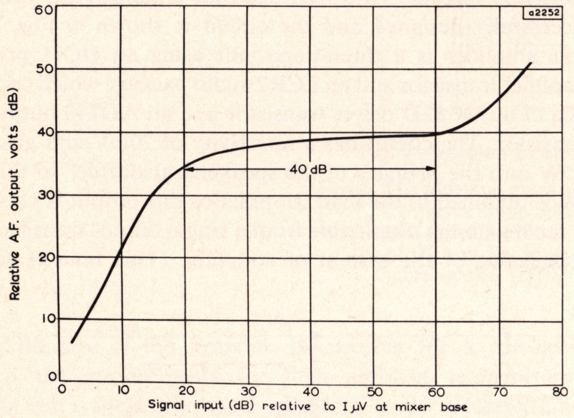

Fig. 2—A.G.C. characteristic

Sensitivity and Signal-to-noise Ratio

A Ferroxcube rod aerial 4in long and 0·375in in diameter is used, type FX1482. The receiver sensitivity is given for three spot frequencies for signal-to-noise ratios greater than 5dB with 5mV at the diode output. This output corresponds to approximately 50mW audio output.

Frequency (Mc/s)	Sensitivity (μV/m)
0·65	250
1·0	100
1·5	100

The sensitivity at 1Mc/s for a 20dB signal-to-noise ratio is approximately 300μV/m.

A.G.C. Characteristic

The a.g.c. characteristic is shown in Fig. 2, where a.f. output at the diode is plotted against r.f. input at the mixer base. For a 6dB change in output, the input can change over a range of 40dB. The maximum signal that the receiver can handle before serious modulation distortion occurs is 200mV/m at 30% modulation. The usable input signal range is 63dB with a.g.c. applied over 40dB.

Selectivity

The i.f. selectivity curve is plotted in Fig. 3 both for a weak signal and a strong signal. The 6dB-bandwidth is ±2·9kc/s and ±2·8kc/s respectively. The attenuation for ±9kc/s is 24dB and 28dB respectively. Approximately 6dB should be added to these figures to obtain the overall r.f. attenuation figures including the aerial circuit attenuation.

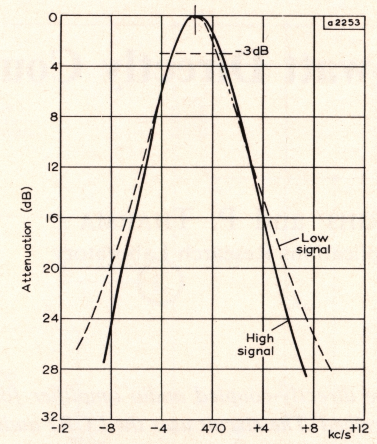

Fig. 3—I.F. selectivity curve

Oscillator

The oscillator voltage across emitter and earth is 95 to 112mV over the medium waveband. Values of minimum output voltage for various r.f. stage line voltages are as follows:

Voltage to r.f. section (V)	Oscillator output (minimum) at emitter (mV)
4·0	95
3·5	73
3·0	53
2·5	30
2·25	20

Below 2·1V, the oscillator ceases to operate, which corresponds to a battery voltage of about 2·6V. The receiver will therefore continue to work when the battery voltage (initially 4·5V) falls by 30% to 3·0V.

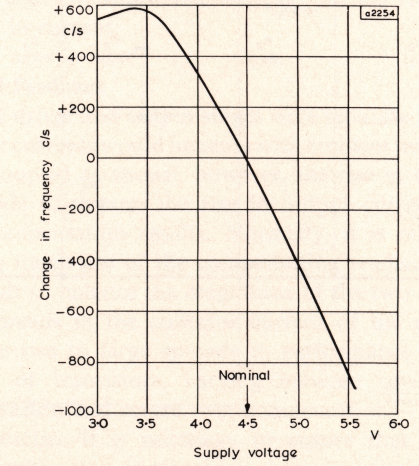

Fig. 4—Variation of oscillator frequency with supply voltage

The variation of the oscillator frequency with supply voltage is shown in Fig. 4 where it can be observed that the change in frequency is 750c/s per V.

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*This circuit is due to P. THARMA. A full description of transformerless audio amplifiers will be the subject of a future article.

Three-watt Directly Coupled Audio Amplifier for Car Radios

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A three-stage directly-coupled audio amplifier for 14V car radios is described. The circuit uses the LCR2 audio package and gives 3W output. The use of the high gain output transistors permits d.c. coupling between driver and output transistors and thus eliminates the conventional driver transformer.

Thermal stability without sacrifice of output power is ensured by a combination of d.c. and thermal feedback.

INTRODUCTION

The output transistor in car radios has so far always been driven by means of a transformer. With the advent of higher gain transistors, however, it has become practicable to consider d.c. coupling between the driver and output transistors. The characteristics of the AD140, with its minimum current amplification factor (h_{fe}) of 30 and low high-temperature collector leakage current, make d.c. coupling particularly attractive.

Within the limitations imposed by the conditions in which a car radio operates, it is very difficult to have a practical design which is economical, has a relatively high output and adequate thermal stability, and accepts practical transistor spreads. However, such an amplifier has been successfully designed and the circuit is shown in Fig. 1. This amplifier is a three-stage unit using an OC71 pre-amplifier transistor and an LCR2 audio package which consists of an OC82D driver transistor and an AD140 output transistor. The circuit has a sensitivity of 20mV and gives 3·5W into the primary of the speaker transformer so that 3W is obtained in the load. In practice this output is close to the maximum obtainable from a single output transistor irrespective of the system of coupling. (This limit is set by the size of the heatsink practicable for a car radio and the requirements for high ambient temperature.) The unit is thermally stable up to 55°C under all conditions, and gives more than 65% of full output at this ambient temperature. Although the majority of amplifiers are stable and give adequate output to well beyond this temperature, there is the danger that the maximum junction temperature may then be exceeded.

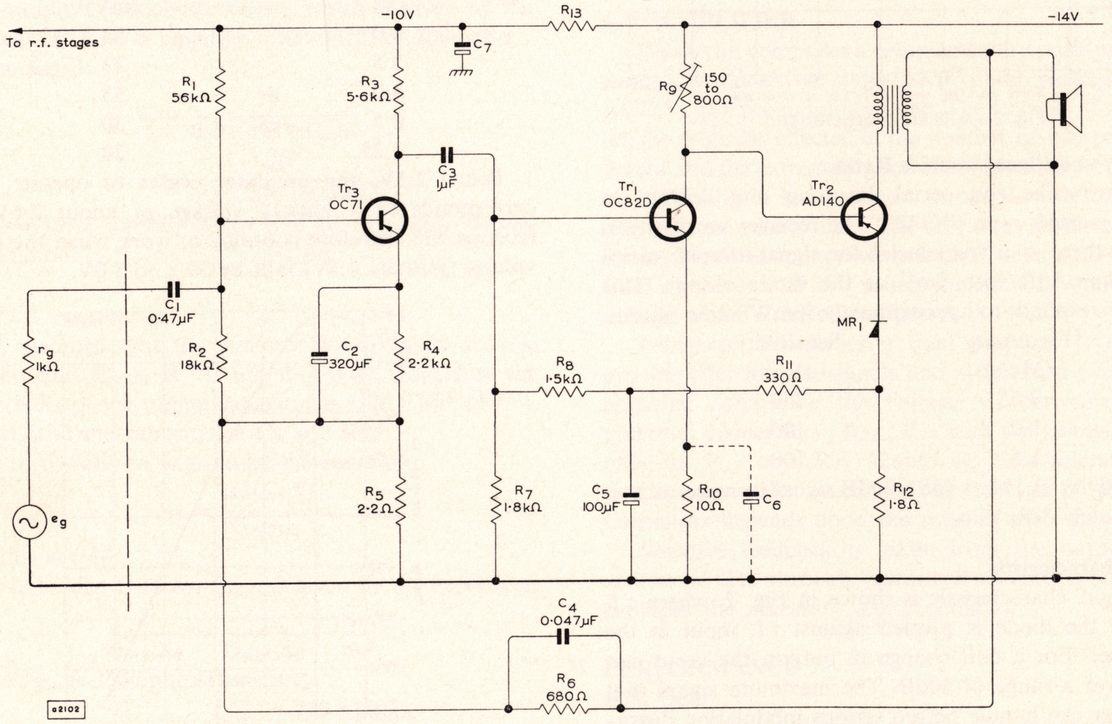

Fig. 1—Version of amplifier circuit with selenium rectifier bias

Note: The use of an extra electrolytic capacitor, that is, capacitor C_6, gives some 7dB extra sensitivity and reduces overall gain spread. If it is used, however, the impedance values of C_4 and R_6 must be doubled.

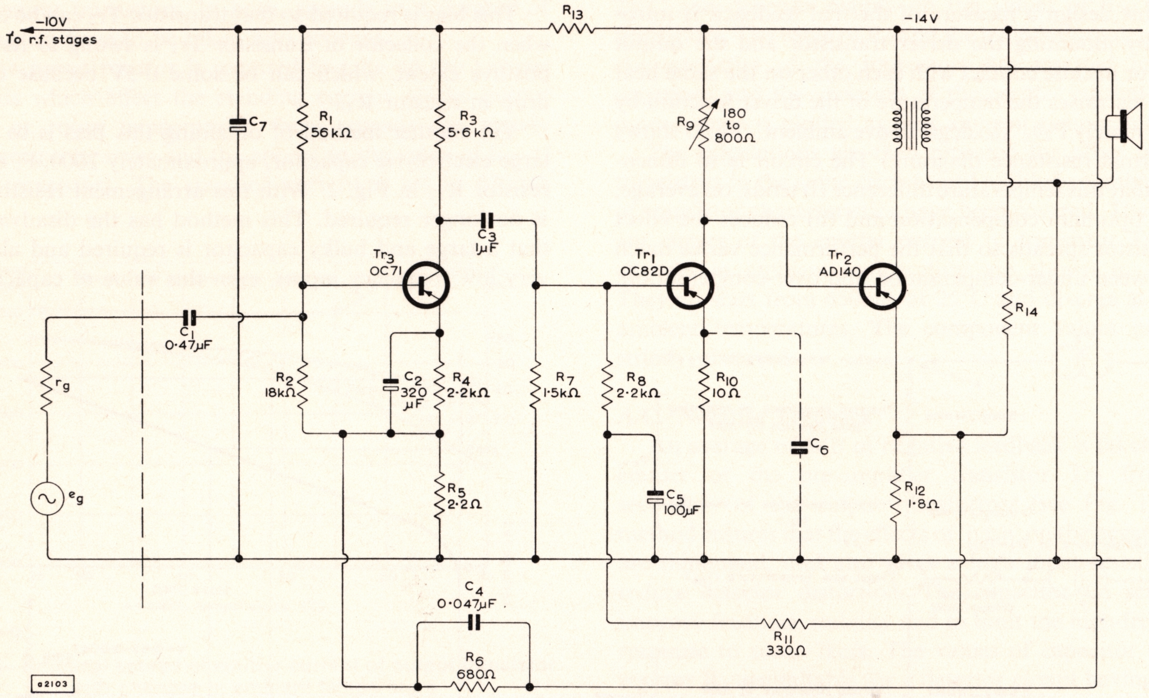

Fig. 2—Version of amplifier circuit with resistive bias

Note: The use of an extra electrolytic capacitor, that is, capacitor C_6, gives some 7dB extra sensitivity and reduces overall gain spread. If it is used, however, the impedance values of C_4 and R_6 must be doubled and the value of R_{14} increased to 100Ω.

The design employs a novel principle, 'thermal feedback', between output and driver transistors to help give the required thermal stability.

The elimination of the bulky and heavy driver transformer used in conventional systems saves space and may make it unnecessary to separate the audio amplifier from the r.f., i.f. and detector units. The saving in cost acquired by not using the transformer will normally more than cover that of the extra low-level transistor stage which is required. One version of the circuit also gives higher gain than is obtainable with the conventional design using a driver transformer.

CIRCUIT DESCRIPTION

The full circuit of the audio unit is shown in Fig. 1 and an alternative version in Fig. 2. The difference between them is only in the means whereby a bias voltage is produced for the emitter of the output transistors.

The pre-amplifier stage of each circuit has its input resistance raised by the injection of negative feedback into its emitter circuit. This stage is followed by the OC82D driver stage which is d.c. coupled to the AD140 output stage. A preset resistor in the collector circuit of the driver transistor Tr_1 is used to set up the collector current of the output stage transistor Tr_2 to its correct value of 630mA. This initial adjustment takes up all the component spreads at room temperature. The speaker is fed from the collector circuit of Tr_2 by the usual output transformer.

A d.c. feedback loop from the emitter of Tr_2 to the base of Tr_1 tends to reduce the current changes caused by changes in temperature. In addition, thermal feedback is used at high ambient temperatures to further reduce these changes in current.

Thermal Feedback

If the driver and output stages were separate, both their collector currents would increase with ambient temperature. When coupled as shown, however, the rise in driver current tends to oppose the rise in output current so that some compensation results. Normally, it is not possible to make much use of the compensating tendency since it is difficult to balance the magnitude of the two influences. Also, spreads in the collector current of the two stages will give rise to large spreads in performance from pair to pair of transistors, varying between severe under-compensation and severe over-compensation.

In practice it is necessary to ensure that the worst likely combination of transistor characteristics and operating conditions will result neither in thermal runaway of the output transistor, nor in so low a current in this transistor that its output capability is unduly reduced at the maximum ambient temperature.

In this design a measure of thermal feedback is introduced by mounting the driver transistor and the output transistor in close contact with each other on the same heat sink. This raises the temperature of the driver junction by approximately $P_2\theta_{y-amb}$degC above ambient. (Fig. 3 shows the thermal resistance diagram.) The circuit is so dimensioned that this temperature difference (i) tends, on average, to give optimum compensation and (ii) reduces the effect of transistor spreads so that the performance varies much less between under-compensation and over-compensation.

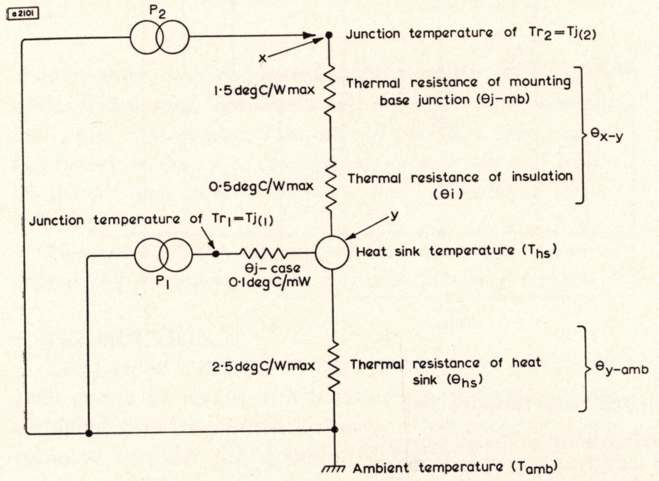
a2101
P2
Junction temperature of Tr2=Tj(2)
x
1·5degC/Wmax
Thermal resistance of mounting base junction (θj-mb)
θx-y
0·5degC/Wmax
Thermal resistance of insulation (θi)
Junction temperature of Tr1=Tj(1)
y
Heat sink temperature (Ths)
P1
θj-case
0·1degC/mW
2·5degC/Wmax
Thermal resistance of heat sink (θhs)
θy-amb
Ambient temperature (Tamb)

Fig. 3—Heat flow diagram

Note: $T_{j(1)} = T_{amb}+P_2\theta_{y-amb}+P_1(\theta_{j-mb}+\theta_{y-amb})$
$\simeq T_{amb}+P_2\theta_{y-amb}$

The second effect is a genuine negative feedback effect: a rise in the output transistor dissipation causes the junction temperature of Tr_1 to rise, so that the emitter current of Tr_1 increases, thereby reducing the emitter current of Tr_2 and opposing the increase in the dissipation of Tr_2.

This, in addition to the effect of the conventional d.c. feedback over the two stages, allows a circuit to be produced which is thermally stable up to an ambient temperature of 60°C and gives at least 2W output up to 55°C.

As the heatsink with its large thermal capacity is an element in the thermal feedback loop, there will be some delay before equilibrium is reached when there is a sudden change in line voltage. Some overshoot will occur but this is of no practical importance except during the initial setting-up. There is no possibility of oscillation because of the thermal feedback system, since there is only one major thermal delay in the feedback loop.

Alternative Versions

The alternative versions of the circuit, shown in Figs. 1 and 2, differ in principle only in the means whereby a bias voltage is produced for the emitter of the output transistor Tr_2.

The circuit shown in Fig. 1 uses a single selenium rectifier in the forward direction to produce a voltage drop of about 0·8V (see Fig. 4) which is only weakly dependent on emitter current. The circuit shown in Fig. 2 uses a 3W resistor, R_{14}, which produces a similar, but slightly lower, bias.

This bias is required so that transistor Tr_2 can be cut-off when the collector of transistor Tr_1 is swung to its most positive extent, which can be some 0·5V because of the drop in resistor R_{10}.

Yet another method of obtaining this bias is to use a large electrolytic capacitor, approximately 2000μF, across resistor R_{12} in Fig. 2. With this arrangement resistor R_{14} is no longer required. This method has the disadvantage that a large and bulky capacitor is required and also, at very low frequency inputs, even this value of capacitance is insufficient to maintain the emitter bias. To avoid clipping under these conditions it is therefore necessary to restrict the low frequency response of the amplifier before the output stage.

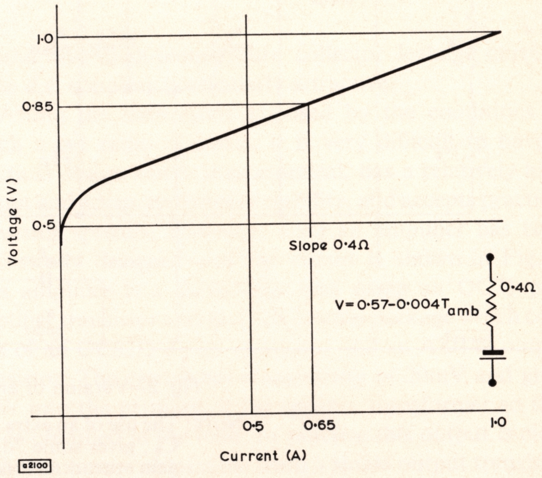

Fig. 4—Forward characteristics and equivalent circuit of selenium rectifier MR_1

The performances of all these versions of the circuit are similar.

A further variant of the circuit is possible and has several advantages, but requires an extra 640μF electrolytic capacitor, C_6, which is shown dotted in Figs. 1 and 2. The use of this capacitor gives some 7dB extra sensitivity and reduces overall gain spreads. Against its use are its cost – approximately 6*d* – and its size, which is ⅜in diameter by 1¼in long. If a lower cut-off frequency of approximately 150c/s is acceptable at the −3dB point, a 320μF, 2·5V capacitor measuring only ⅜in diameter by ⅞in long is sufficient.

When a capacitor such as C_6 is used, the impedance of the a.c. feedback components R_6 and C_4 has to be doubled to maintain the feedback value. In the circuit shown in Fig. 2 it is also possible to reduce the amount of bias on the emitter of Tr_2 by increasing the value of R_{14} and thus reducing the power dissipated in it. Changes in the values of R_7 and R_8 are then also required.

CIRCUIT DESIGN

The difficulty of finding a practical design for the d.c. coupled pair of transistors is entirely one of choosing circuit values which will ensure satisfactory operation over the full temperature range for all likely combinations of driver

and output transistors. In both, spreads in a number of parameters have to be taken into account.

One preset resistor is required to take up the various spreads which affect the standing current of the output stage at room temperature; these spreads are mainly resistance tolerances and the spread in the value of h_{fe} of both transistors. The standing current is set to 630mA which gives a quiescent dissipation of 8W for a 14V line voltage and 10W at 15·7V. The nominal output is 3·5W into a collector load of 18Ω.

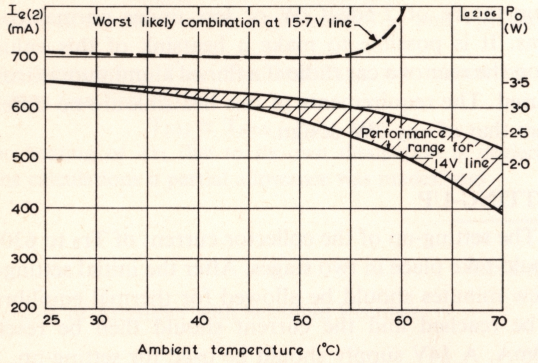

Fig. 5—Practical spread of emitter current of output transistor with changes in ambient temperature

Driver Stage

In order to obtain the thermal compensating action, the driver stage has to be designed with a certain relatively low degree of d.c. stability. Although this results in a considerable spread in stage current (10 to 28mA), it is turned to advantage in the design. Normally the driver current must always be somewhat higher than the maximum base current swing of the output transistor. With the LCR2 package, however, it is not necessary to design the minimum possible driver current to be high enough to drive an output transistor having the lowest possible value of current amplification factor.

By virtue of cross matching in the package, transistors having low values of h_{fe} will always be accompanied by transistors having high values of h_{fe} which automatically set themselves up in the circuit to pass relatively high drive current. The driver current in any pair will thus be as low, apart from a margin, as is permissible for its output partner. This fact is also used in the calculation of the range required for the preset variable resistor.

Thermal Stability

Because d.c. and thermal feedback are used in the circuit, the equations which govern the circuit behaviour with changes in temperature are very complex. The simultaneous non-linear equations which describe the variation of the stage currents with ambient temperature are given in the Appendix.

The following steps are required in the calculation for any one set of circuit values:

The value of the preset resistor R_9 for any one pair of transistors is calculated from the circuit equations at 25°C and depends mainly on the values of h_{fe} for the transistors. The value of R_9 and the temperature-dependent characteristics of the transistors are inserted into the equations and the variation of the emitter current of the output stage is plotted against ambient temperature.

The calculations are then repeated for other pairs of transistors in the same circuit until all likely combinations of transistor characteristics have been covered. The resultant curves of the emitter current of transistor Tr_2 with changes in ambient temperature are shown in Fig. 5. They all start from 650mA at 25°C and diverge at higher ambient temperatures. The component values must be strictly observed.

A.C. Negative Feedback

An average of 6dB of negative feedback is taken to the emitter of the pre-amplifier transistor Tr_3 from the secondary of the output transformer (see Fig. 1). This method ensures that the effects of 'hash' on the negative line are minimised, and also gives a high input resistance to reduce detector distortion. The capacitor C_4 serves to improve feedback stability and to limit the high-frequency response to some 7kc/s. The values of capacitor C_4 and resistor R_6 are suitable for a detector circuit having a 1kΩ source impedance, that is, one with a 2·5kΩ volume control.

PERFORMANCE

Output Power

The maximum output power into the transformer primary is 3·5W.

Sensitivity

Sensitivity values are given in Table 1 in terms of the e.m.f. required in a source of 1kΩ resistance (to simulate a detector circuit with a 2·5kΩ control) with and without capacitor C_6.

TABLE 1
Sensitivity
(Source impedance = 1kΩ)

Input with C_6	Input without C_6	Power output
1·1mV±2dB	2·3mV±3dB	50mW
9mV ±3dB	20mV ±3dB	3·5W

Distortion

The average distortion at 3·5W output is 5%.

Input Resistance

The typical input resistance is 4kΩ, with 2kΩ being the minimum.

Behaviour with Ambient Temperature Changes

Fig. 5 shows the practical limits of the spread in performance which can be expected with changes in ambient temperature. Less than 1 in 5000 LCR2 packages will fall outside these limits.

At the nominal line voltage, all combinations of transistors are safe from thermal runaway at any temperature.

Moreover, the emitter current of the output transistor will be greater than 510mA, and thus gives more than 2W output at 55°C ambient temperature.

Under sustained high voltage conditions (15·7V) and for the worst likely combination, the output transistor is thermally stable up to 60°C. It should be noted, however, that for this combination, the absolute maximum junction temperature of Tr_2 which is 100°C, is reached at 56°C ambient temperature.

At this temperature the dissipation is $(15{\cdot}7-1{\cdot}8)\left(720\times\frac{49}{50}\right)$W, or 9·9W, and the junction temperature is $56+(4{\cdot}5\times9{\cdot}9)$ or 100°C.

The majority of output transistors will show decreasing values of emitter current with ambient temperature characteristics even under sustained high line voltage conditions. They will thus be safe from thermal runaway at any temperature.

High line voltage conditions of a duration less than the warm-up time of the heatsink are of no practical importance.

CONSTRUCTIONAL DETAILS

As the design of the circuit depends largely on thermal feedback it is essential that the driver transistor is thermally closely coupled to the output transistor. This means that both transistors must be firmly mounted close to each other on the same heatsink. Silicone grease should be used on each side of the mica washers.

The heatsink must have a thermal resistance to ambient temperature of 2·5degC/W or less under normal conditions. It is possible to make a heatsink of this value to fit on the rear of a car radio if a finned aluminium extrusion is used. The requirements of this heatsink are no different from that of any 3W design.

SETTING-UP

The setting-up of the collector current of Tr_2 to 630mA should take place in two stages. After the initial setting-up, a few minutes should be allowed for thermal equilibrium to be reached and the current should then be reset to 630mA. A 14V supply should be used for setting-up.

APPENDIX

Outline of Thermal Stability Calculation

The equations relating to the temperature stability of the directly-coupled pair are given in this appendix. The d.c. circuits of the two versions of the directly-coupled pair are shown in Figs. 6 and 7.

The selenium rectifier shown in Fig. 6 has been considered as a composite unit with transistor Tr_2. The bias voltage V_3 in Fig. 7 is the voltage produced by the current flow from R_{14} across R_{12} in Fig. 2.

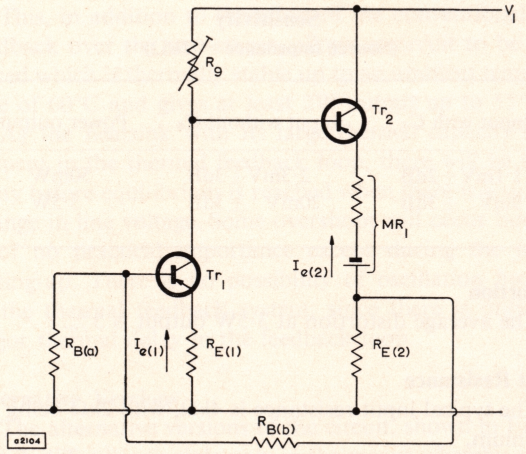

Fig. 6—Equivalent d.c. circuit of directly coupled pair with selenium rectifier

Approximate circuit values were established and these were used in a computer programme to find realistic performance spreads. The actual characteristics of some OC82D and AD140 transistors were also inserted into the programme. The resultant curves of the emitter current values of transistor Tr_2 with changes in ambient temperature were checked against experimental results from a temperature run to verify the calculations. Agreement was satisfactory.

The following equations use the standard symbols for the transistor characteristics with suffixes 1 and 2 to denote transistors Tr_1 and Tr_2 respectively.

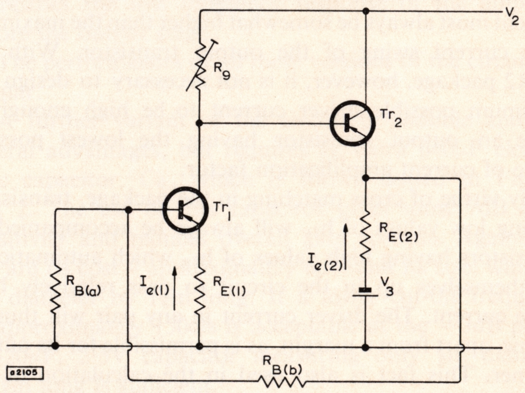

Fig. 7—Equivalent d.c. circuit of directly coupled pair with resistive bias

For the circuit shown in Fig. 7:

$$I_{e(1)} = G_1\left[\frac{1}{R_{B(b)}}(I_{e(2)}R_{E(2)}+V_3) - \frac{V_{BE(1)}}{mR_{B(b)}} + I_{CBO(1)}\right],$$

and

$$I_{e(2)} = G_2\left[\frac{1}{R_9}(V_1-V_3-V_{BE(2)})+I_{CBO(2)}\right]-I_{e(1)}G_2\,;$$

where

$$m = \frac{R_{B(a)}}{R_{B(a)}+R_{B(b)}},$$

$$G_1 = \frac{1}{(R_{E(1)}/mR_{B(b)})+1/h_{fe(1)}},$$

and

$$G_2 = \frac{1}{(R_{E(2)}/R_9)+1/h_{fe(2)}}.$$

$V_{BE(1)}$, $V_{BE(2)}$, $I_{CBO(1)}$ and $I_{CBO(2)}$ are functions of temperature as given below:

$$V_{BE(1)} = A-2.10^{-3}T_{j(1)}V,$$
$$V_{BE(2)} = B-2.10^{-3}T_{j(2)}V,$$
$$I_{CBO(1)} = D_{e0\cdot0935}T_{j(1)}\mu A,$$

and

$$I_{CBO(2)} = F_e^{0\cdot0935}T_{j(2)}\mu A,$$

where A, B, D, and F are constants and $T_{(j)1}$ and $T_{j(2)}$ are in degC.

The temperatures $T_{j(1)}$, $T_{j(2)}$ and T_{amb} are related by the equations (see Fig. 3):

$$T_{j(1)} \simeq T_{amb}+P_2\theta_{y-amb},$$

if the heating of Tr_1 due to its own dissipation is neglected (in the calculations a partial allowance was made), and

$$T_{j(2)} = T_{amb}+P_2(\theta_{x-y}+\theta_{y-amb})$$

where

$$P_2 = I_{e(2)}V_{ce}.$$

For the selenium rectifier version (see Fig. 6):

$$V_3 = 0,$$
$$V_{BE(2)} = B+0{\cdot}85-2{\cdot}10^{-3}T_{j(2)}-4{\cdot}10^{-3}T_{amb}$$

as the voltage drop across the rectifier is 0·85V at 0·65A and its temperature coefficient is −4mV/degC.

The above set of simultaneous equations are of the form:

$$I_{e(1)} = f(I_{e(2)}, T_{j(1)}),$$
$$I_{e(2)} = f_1(I_{e(1)}, T_{j(2)}),$$
$$T_{j(1)} = f_2(T_{amb}, I_{e(2)})$$

neglecting the contribution due to self-heating which involves $I_{e(1)}$ also, and

$$T_{j(2)} = f_3(T_{amb}, I_{e(2)}).$$

For the selenium rectifier version $I_{e(2)}$ is also directly a function of T_{amb}.

LIST OF COMPONENTS

a) Version of Amplifier with Selenium Rectifier

Resistors

Circuit Ref.	Value	Tolerance (±%)	Rating (W)
R_1	56 kΩ	10	¼
R_2	18 kΩ	10	¼
R_3	5·6 kΩ	10	¼
R_4	2·2 kΩ	10	¼
R_5	2·2 Ω	10	¼
R_6	680 Ω (1·5kΩ when C_6 is used)	10	¼
R_7	1·8 kΩ	10	¼
R_8	1·5 kΩ	10	¼
R_9	150 to 800Ω (preset)	10	¼
R_{10}	10 Ω	10	¼
R_{11}	330 Ω	10	¼
R_{12}	1·8 Ω	10	1
R_{13}	Depends on r.f. circuit		

Capacitors

Circuit Ref.	Value	Mullard type number	Rating (V)
C_1	0·47 µF	C296AA/A470K	125
C_2	320 µF	C426AE/A320	2·5
C_3	1 µF	C296AA/A1M	125
C_4	0·047µF	C296AA/A47K	125
	or, when C_6 is used		
	0·022µF	C296AA/A22K	125
C_5	100 µF	C426AE/B100	4·0
C_6	320 µF	C426AE/A320	2·5
	or		
	640 µF	C426AE/A640	2·5

Transistors

Circuit Ref.	Mullard type number	
Tr_1	OC82D	LCR2 package
Tr_2	AD140	
Tr_3	OC71	

For both versions, feedback components are given for a 3Ω loudspeaker. The transformer ratio should be chosen to provide 18Ω collector load after allowing for winding resistances.

(b) Version of Amplifier with Resistive Bias

Resistors

Circuit Ref.	Value	Tolerance (±%)	Rating (W)
R_1	56 kΩ	10	¼
R_2	18 kΩ	10	¼
R_3	5·6 kΩ	10	¼
R_4	2·2 kΩ	10	¼
R_5	2·2 Ω	10	¼
R_6	680 Ω (1·5kΩ when C_6 is used)	10	¼
R_7	1·5 kΩ	10	¼
R_8	2·2 kΩ (1·8kΩ when C_6 is used)	10	¼
R_9	150 to 800Ω (preset)	10	¼
R_{10}	10 Ω	10	¼
R_{11}	330 Ω	10	¼
R_{12}	1·8 Ω	10	1
R_{13}	Depends on r.f. circuit		
R_{14}	56 Ω	10	3
	or, when C_6 is used		
	100 Ω	10	2

Capacitors

Circuit Ref.	Value	Mullard type number	Rating (V)
C_1	0·47 µF	C296AA/A470K	125
C_2	320 µF	C426AE/A320	2·5
C_3	1 µF	C296AA/A1M	125
C_4	0·047µF	C296AA/A47K	125
	or, when C_6 is used		
	0·022µF	C296AA/A22K	125
C_5	100 µF	C426AE/B100	4·0
C_6	320 µF	C426AE/A320	2·5
	or		
	640 µF	C426AE/A640	2·5

Transistors

Circuit Ref.	Mullard type number	
Tr_1	OC82D	LCR2 package
Tr_2	AD140	
Tr_3	OC71	

Direct-coupled Amplifiers for Hearing Aids

In the article which we published in the March 1962 issue of Mullard Technical Communications we specified a microphone and an earpiece which were suitable for the circuits. They were referred to by their development numbers.

We learn from the manufacturer—Knowles Electronics Ltd.—that these components are now marketed under production type-numbers, as follows:

Microphone:	**BA1501**	(formerly BA1481)
Earpiece:	**BB1511**	(formerly BB1491)

These components are available from Knowles Electronics Ltd. at Victoria Road, Burgess Hill, Sussex.

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Corrections

Vol. 6, No. 60, page 409
In the ninth line of the right-hand column, read *conductance* in place of *susceptance.*

Vol. 7, No. 62, page 60
In the fourth line from the bottom of the right-hand column the assumed value of C_{OUT} should be b_{oe}/ω, not B_{oe}/ω.

Vol. 7, No. 62, page 63
The left-hand side of the expression immediately below Fig. 12 should be Y_{ie}, not V_{ie}.

The three expressions preceding (A8) should be introduced by the words *If . . . and . . . then . . .* The existing wording (*If . . . then . . . so that . . .*) is incorrect.